



Research Department Report

December 1987

**The relationship between
vision carrier-to-noise ratio and
picture signal-to-noise ratio in a
System I television receiver**

A.P. Robinson, B.Sc., A.R.C.S.

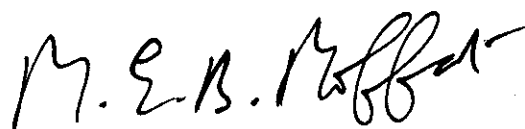
**THE RELATIONSHIP BETWEEN VISION CARRIER-TO-NOISE RATIO AND
PICTURE SIGNAL-TO-NOISE RATIO IN A SYSTEM I TELEVISION RECEIVER**

A.P. Robinson, B.Sc., A.R.C.S.

Summary

Definitions of vision carrier-to-noise ratio and picture signal-to-noise ratio are given, including a definition of noise bandwidth appropriate to a vestigial-sideband receiver. For television System I, it is shown both by theory and experiment that the picture signal-to-noise ratio is 8 dB less than the vision carrier-to-noise ratio. The effect on this result of filtering in the system is discussed and the noise characteristics of a domestic television receiver are presented. A justification is given for the minimum field strengths recommended by the CCIR for a television service.

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1. INTRODUCTION

Many factors go into the planning of a network of transmitters. The broadcaster's aim is that the viewer should obtain satisfactory picture quality. Generally, the field strength of the signal at the viewer's location is regarded as the key to picture quality.

Let us consider impairments to picture quality caused by noise. As the signal power applied to a receiver is decreased, so the level of noise displayed on the picture increases. To relate the field strength of the signal to the picture signal-to-noise ratio, several pieces of information are required:

- (i) the gain of the aerial system,
 - (ii) the noise temperature of the receiving installation,
 - (iii) the noise bandwidth of the receiver,
 - (iv) the relationship between the vision carrier-to-noise ratio (C/N) and the picture signal-to-noise ratio (S/N),
- and (v) the baseband noise spectrum that is produced, if the weighted S/N is required.

This Report concentrates on the aspects connected with the receiver. The question of relating the carrier power and the noise power in the r.f. channel to the signal power and the noise power after the demodulator often presents difficulties. The problem is particularly awkward in a television system because it requires an analysis of the operation of a demodulator in a vestigial sideband (v.s.b.) system and C/N and S/N are defined in different ways.

The relationship between C/N and S/N has been considered by other authors^{1,2,3} but some of the accounts are unclear, and the conclusions conflict. In this Report the question is examined both by theory and experiment.

2. DEFINITION OF C/N AND S/N

The carrier-to-noise ratio may be thought of as being measured at the input to the receiver. In such a measurement, the relevant noise level is the equivalent input noise level. (The derivation of C/N from the field strength and the parameters of the receiving

installation is given in Section 3.4.) The carrier-to-noise ratio may also be measured at any point in the receiver which comes after significant sources of noise and before significant filtering (e.g. in many television receivers, the tuner output).

The carrier power is defined as the mean power of the vision carrier during the synchronising pulse ('peak sync power').

The noise is white and wideband at the point where it is measured, therefore it is necessary to define a bandwidth in which the mean noise power is to be measured. For a standardised measurement, the most appropriate noise bandwidth to use is that of the ideal receiver filter. The idealised response is shown in Fig. 1. The noise bandwidth of a filter is the width of an assumed rectangular filter passing the same mean noise power (with white noise incident at the same level). The noise bandwidth, B , can be expressed mathematically in terms of the frequency response of the filter, $H(f)$:

$$B = \int |H(f)|^2 df$$

Hence the noise bandwidth of the ideal receiver filter for System I is 5.08 MHz. The value of C/N referred to this bandwidth will be denoted by $(C/N)_0$. It may be noted that other vestigial sideband responses than that shown in Fig. 1 are possible. The same demodulated video signal will still be obtained, provided that the filter response is skew symmetrical about the vision carrier frequency, and reaches zero and unity within 1.25 MHz of the vision carrier frequency. One possibility is a cosine roll-off, although the filter would have a noise bandwidth greater than 5.08 MHz. The optimum (idealised) response from the point of view of noise performance is shown in Fig. 2. The noise bandwidth is only 4.88 MHz; however it would be impractical to implement such a response.

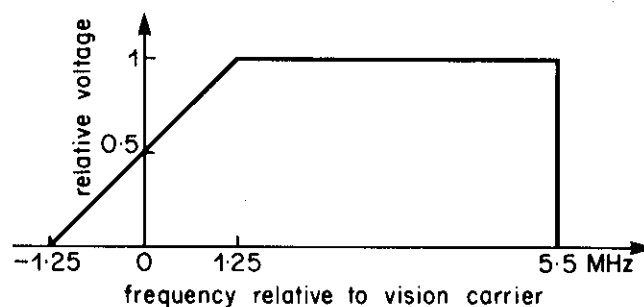


Fig. 1 - Idealised television receiver i.f. filter response (without spectrum inversion).

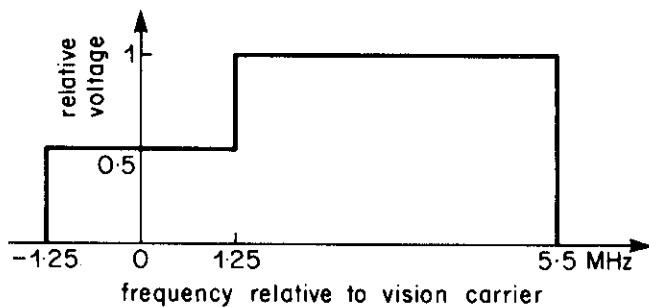


Fig. 2 - Idealised television receiver i.f. filter response, optimised for noise performance.

Alternatively the noise bandwidth of the receiver in question may be used* and the result measured in this way will be denoted by $(C/N)_{rec}$.

The above definition of noise bandwidth does not specify how to deal with a filter which has a non-zero insertion loss, or which is preceded by an amplifier. In such cases it is necessary to specify a reference level, i.e. the 'height' of the rectangular filter, relative to the response of the actual filter or filter-amplifier combination. In many applications the reference level chosen is the level of the peak of the response and often this coincides with the carrier frequency of the system under consideration. This approach achieves a simple connection between the measurements of carrier level and noise level, independent of the exact shape of the filter response. This idea can be extended to System I (v.s.b.) receiver filters, by specifying the reference level to be 6 dB above the level of the response at the vision carrier frequency. The noise bandwidth of the ideal filter by this definition is still 5.08 MHz. It will be seen that this definition makes the relationship between $(C/N)_{rec}$ and S/N virtually independent of the receiver.

The signal-to-noise ratio is measured on the video waveform. It is defined as the ratio of peak-to-peak picture voltage (usually 0.7 V) to r.m.s. noise voltage. For simplicity, a slightly non-standard definition is adopted: it is assumed that there is no significant filtering in the measuring instrument and that the receiver satisfactorily defines the bandwidth of the baseband noise measurement. The standard practice would be to make measurements through a 5 MHz low pass filter.† Such a filter would remove some of the noise and increase the measured value of the signal-to-noise ratio. This is considered further in Section 3.3.

* A practical receiver always has baseband filtering, e.g. to remove the sound carrier and i.f. components from the demodulator output. The baseband filtering can be converted to an equivalent at r.f. and then the overall noise bandwidth can be found.

† The purpose of the low pass filter is to eliminate spurious and out-of-band components and the choice of 5 MHz bandwidth makes the measurement applicable to systems with a video bandwidth of 5 or 5.5 MHz. A typical instrument can also introduce various other weighting and band-limiting filters.

3. THEORETICAL RELATIONSHIP

To derive a relationship between the carrier-to-noise ratio and the signal-to-noise ratio we need to relate the picture level to the carrier power, and the baseband noise voltage to the r.f. noise level. The peak-to-peak picture level can be derived from the peak sync carrier power, knowing the modulation parameters, and taking into account the v.s.b. characteristic. In deriving the baseband noise voltage, account must be taken of the r.f. filtering and also of any baseband filtering.

A receiver can be simplified to the notional block diagram of Fig. 3 for the purpose of analysis. It is assumed that the noise is white prior to any filtering in the receiver and that significant sources of noise come before the filtering that defines the receiver's frequency response. Then all the filtering in the receiver can be attributed to a single equivalent filter at i.f. (referred to as 'the filter'). A synchronous demodulator is assumed, although an envelope detector gives similar results except when the noise level is high. (Most receivers of recent design have a synchronous demodulator.) The block diagram also shows many of the symbols used in the following analysis to denote the levels at various points.

3.1 Relationship between carrier power and peak-to-peak picture level

The standard levels of a video signal are shown in Fig. 4 and the standard levels of the modulated vision carrier are shown in Fig. 5. Suppose the active line contains a low frequency tone of 0.7 V peak-to-peak. Then from inspection of Fig. 5, during the active line the carrier component has an amplitude of 48% of peak sync and the two sideband components each have an amplitude of 14%.

Referring to Fig. 3, let

C = peak sync carrier power incident on the filter;

C_{IF} = peak sync carrier power emerging from the filter;

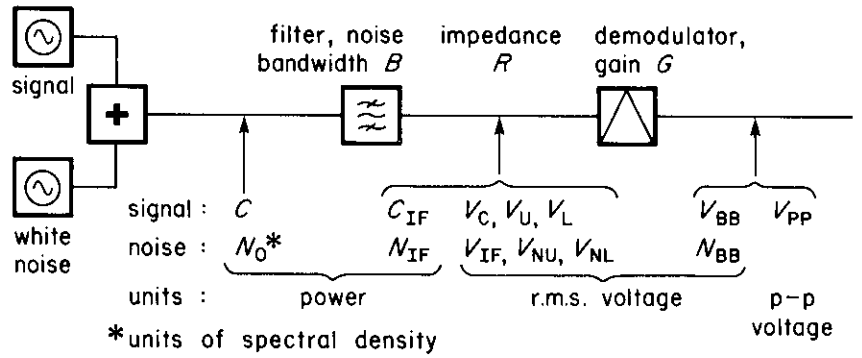
R = system impedance;

V_C = peak sync carrier voltage (r.m.s.) after the filter;

V_U = r.m.s. voltage of upper side tone after the filter corresponding to a full amplitude low frequency video tone;

V_L = r.m.s. voltage of lower side tone after the filter corresponding to a full amplitude low frequency video tone.

Fig. 3
Block diagram of system for theoretical analysis, showing the symbols used in the text to denote the levels of signal and noise at each point, and the respective units of measurement.



For consistency with the definition of noise bandwidth in Section 2, let the filter response be scaled so that the insertion loss at the vision carrier frequency is 6 dB. Then the carrier power and voltage after the filter are

$$C_{IF} = 0.25 C \quad (1)$$

$$V_C = \sqrt{RC_{IF}}$$

$$= 0.5 \sqrt{RC} \text{ from (1)} \quad (2)$$

If the filter response is skew symmetrical in the vicinity of the vision carrier frequency then the sideband amplitudes after the filter are

$$V_U = 0.14 (1 + \Delta) V_C$$

(the transmitted sideband amplitude is 14% of peak sync.)

$$= 0.07 (1 + \Delta) \sqrt{RC} \text{ from (2)} \quad (3)$$

Similarly,

$$V_L = 0.07 (1 - \Delta) \sqrt{RC} \quad (4)$$

where Δ is a small quantity representing the amount of skew.

Let V_{BB} = r.m.s. voltage of full amplitude low frequency video tone after demodulator;

V_{PP} = peak-to-peak voltage of full amplitude low frequency video tone after demodulator, i.e. peak-to-peak picture voltage;

G = voltage gain of demodulator.

The amplitudes of coherent sidebands add in the demodulator, producing a baseband video level of

$$V_{BB} = G (V_U + V_L)$$

$$= G \times 0.07 (1 + \Delta + 1 - \Delta) \sqrt{RC} \text{ from (3) and (4)} \quad (5)$$

Hence the peak-to-peak picture voltage is

$$V_{PP} = 2 \sqrt{2} V_{BB}$$

$$= 2 \sqrt{2} G \times 0.14 \sqrt{RC} \text{ from (5).} \quad (6)$$

3.2 Relationship between r.f. noise level and baseband noise voltage

Let N_0 = noise power spectral density incident on the filter;

N_{IF} = noise power emerging from the filter;

V_{IF} = r.m.s. noise voltage after the filter;

V_{NU} = noise voltage in upper sideband after the filter;

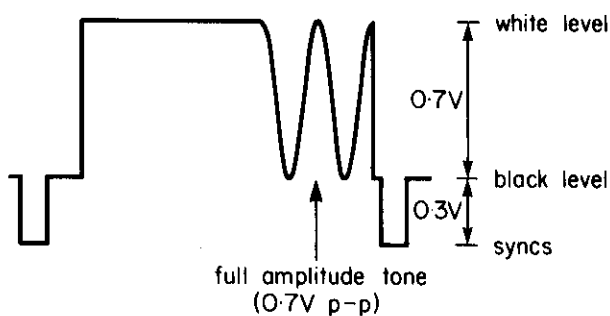


Fig. 4 - Standard levels of video.

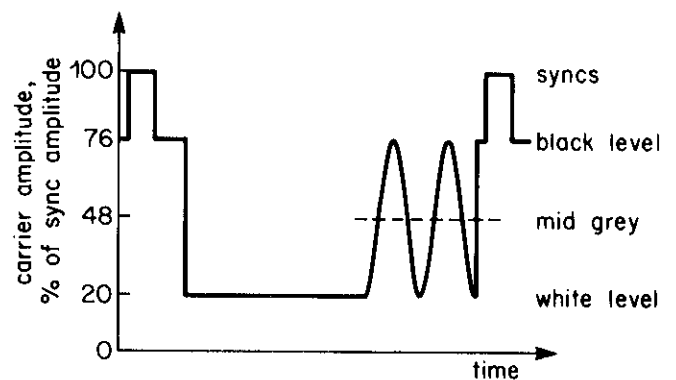


Fig. 5 - Standard levels of vision carrier (for low video frequencies).

V_{NL} = noise voltage in lower sideband after the filter;

N_{BB} = r.m.s. noise voltage after demodulator.

The noise voltage after the filter can be related to the r.f. noise density as follows:

$$\begin{aligned} N_{IF} &= N_o B \\ V_{IF} &= \sqrt{RN_{IF}} \end{aligned} \quad (7)$$

We can consider the noise power after the filter to be split between the upper and lower sidebands. Bearing in mind that power is proportional to the square of voltage, we can relate the voltages in the two sidebands to the total voltage:

$$V_{NU}^2 + V_{NL}^2 = V_{IF}^2 \quad (8)$$

After the demodulator, the two sidebands give baseband noise voltages which are respectively

$$GV_{NU} \text{ and } GV_{NL}$$

The two contributions are incoherent so they undergo power addition, thus the total baseband noise voltage is given by

$$N_{BB}^2 = (GV_{NU})^2 + (GV_{NL})^2$$

Hence

$$\begin{aligned} N_{BB} &= GV_{IF} \quad \text{from (8)} \\ &= G \sqrt{RN_{IF}} \quad \text{from (7)}. \end{aligned} \quad (9)$$

3.3 Relationship between C/N and S/N

From the definition, the carrier-to-noise ratio is

$$(C/N)_{rec} = \frac{C}{N_{IF}} \quad (10)$$

The definition of signal-to-noise ratio gives

$$S/N = \frac{V_{PP}}{N_{BB}}$$

We can now connect the two ratios

$$\begin{aligned} S/N &= \frac{0.28 \sqrt{2} G \sqrt{RC}}{G \sqrt{RN_{IF}}} \quad \text{from (6) and (9)} \\ &= 0.28 \sqrt{2} \sqrt{\frac{C}{N_{IF}}} \end{aligned}$$

$$= 0.28 \sqrt{2} \sqrt{(C/N)_{rec}} \quad \text{from (10)}$$

or in decibels

$$\begin{aligned} S/N \text{ (dB)} &= (C/N)_{rec} \text{ (dB)} + 20 \log_{10}(0.28 \sqrt{2}) \\ S/N \text{ (dB)} &= (C/N)_{rec} - 8.0 \text{ dB}. \end{aligned} \quad (11)$$

It should be noted that no constraint has been imposed upon the response of the filter except that it should be skew symmetrical in the vicinity of the vision carrier frequency. This constraint is likely to be satisfied in any practical receiver.

It may be more useful to have the result expressed in terms of the standardised value of C/N . From the definitions

$$(C/N)_{rec} = (C/N)_o \times \frac{5.08 \times 10^6}{B}$$

Hence

$$\begin{aligned} S/N &= (C/N)_o + 10 \log_{10} \left(\frac{5.08 \times 10^6}{B} \right) \\ &\quad - 8.0 \text{ dB} \quad \text{from (11)}. \end{aligned}$$

It is also possible to take account of a 5 MHz low pass filter typically introduced in a video noise meter. The expression becomes

$$\begin{aligned} S/N &= (C/N)_o + 10 \log_{10} \left(\frac{5.08 \times 10^6}{B} \right) \\ &\quad + x - 8.0 \text{ dB} \end{aligned}$$

The factor x is positive and it allows for the noise removed by the low pass filter. The value of x depends on the responses of the receiver filter and the 5 MHz low pass filter. With idealised filtering,

$$x = 10 \log_{10} (5.08/4.58) = 0.4 \text{ dB}.$$

Using the filter recommended in CCIR Rec. 567, the value of x is 0.6 dB (still assuming the idealised i.f. filter in the receiver).

3.4 Extension of relationship to include connection with field strength

In the Introduction, various parameters of the receiving installation are mentioned. By taking these into account, the carrier-to-noise ratio, and hence the signal-to-noise ratio, can be related to the field strength. From Appendix 1, the relationship between carrier-to-noise ratio and field strength is

$$(C/N)_{\text{rec}} = E \text{ (dB } \mu\text{V/m)} + 20\log \lambda + G - L \\ - 10\log k_B T_o B F - 154.6 \text{ dB}$$

where E = r.m.s. electric field strength (peak sync);

λ = wavelength, m;

G = gain of aerial relative to a dipole, dB;

L = feeder loss, dB (> 0);

k_B = Boltzmann's constant ($1.38 \times 10^{-23} \text{ JK}^{-1}$);

T_o = reference temperature (290 K);

F = system noise factor referred to receiver input.

Hence the relationship between signal-to-noise ratio and field strength is

$$S/N = E \text{ (dB } \mu\text{V/m)} + 20\log \lambda + G - L \\ - 10\log k_B T_o B F + x - 162.6 \text{ dB}$$

Using this equation, the CCIR planning limits for field strength⁴ can be explained, as shown in Appendix 1.

4. EXPERIMENTAL RELATIONSHIP

4.1 An experiment with a simplified system

The first experiment was performed to verify the principles using the simplest representative configuration. The key elements are the filter and the demodulator.

A block diagram of the experiment is shown in Fig. 6. The experiment was carried out at i.f., using a passive filter. A known amount of noise was added to a vision signal. The signal was then passed through the filter, and a synchronous demodulator, and the video was amplified to standard level. The resulting noise was measured with a power meter.

The frequency response of the filter, shown in Fig. 7, was a good approximation to the ideal. The response was measured with a synthesised sweeper, and the data was captured by a computer which was programmed to calculate the noise bandwidth, which was 5.21 MHz. The associated experimental error is believed to be less than ± 0.1 dB.

The spectrum of the noise source was found to be flat to within 0.1 dB over the frequency range of interest. The measured noise power at the output of the filter was $1.60 \mu\text{W}$: this is, in effect, the noise power incident on the filter, measured in the noise bandwidth of the filter. The modulation depth of the signal was set with an estimated error of 0.15 dB and the power of a plain carrier at the peak sync level was found to be $960 \mu\text{W}$ at the output of the filter. The level incident on the filter would be 6.02 dB higher, and thus $(C/N)_{\text{rec}}$ was 33.80 dB.* The insertion loss of the filter does not need to be taken into account because it cancels out in the calculation of $(C/N)_{\text{rec}}$. The uncertainty in $(C/N)_{\text{rec}}$ is 0.2 dB; this includes the instrumentation uncertainty associated with the measurement of the ratio $960 \mu\text{W}/1.60 \mu\text{W}$ and for convenience it also includes the experimental error in the relationship between the measured peak sync level and the difference between black level and white level.

* Two decimal places are retained to avoid the accumulation of rounding errors although the second decimal place is of no significance.

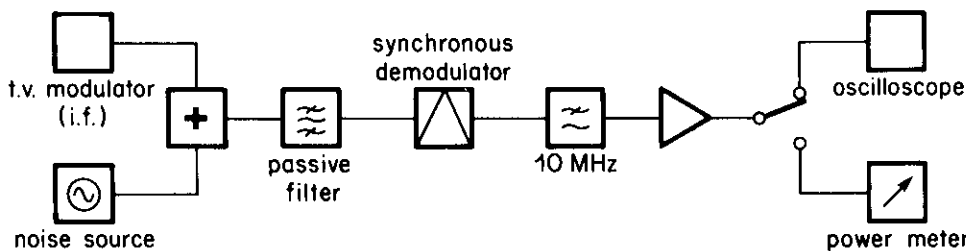


Fig. 6
Block diagram of
first experiment.

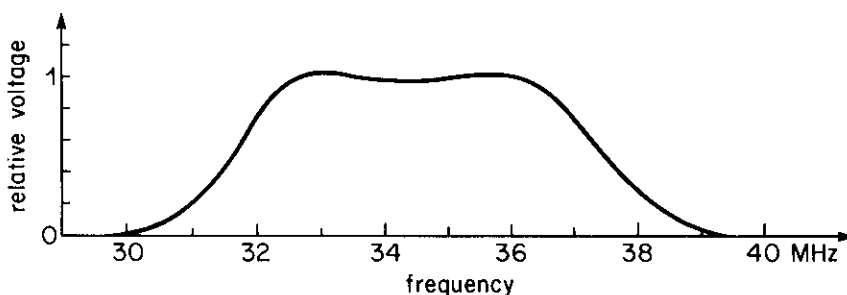


Fig. 7 - Frequency response of passive
filter (vision i.f. = 37.5 MHz).

The peak-to-peak picture level was set to 0.7 V (into 75 ohms) with a calibrated oscilloscope; the experimental error in this adjustment was estimated to be 0.15 dB. With plain carrier at i.f., the baseband noise power was 15.6 μ W; the uncertainty in this measurement is thought to be 0.1 dB. The linearity of the system was shown to be good. Thus S/N was 26.22 dB with an uncertainty of 0.2 dB.

Hence the ratio between $(C/N)_{rec}$ and S/N was 7.58 dB with an uncertainty of 0.3 dB, where all the errors have been combined by the root-sum-of-squares. The estimated error bounds of the experimental result just fail to encompass the theoretical value of 8.0 dB of equation (11). This may be accounted for by other sources of error such as the behaviour of the demodulator, the baseband filtering and impedance mismatches.

The spectrum of the baseband noise is shown in Fig. 8, together with the overall (video) frequency response of the system. In the region between 0 and 1.25 MHz where both sidebands contribute, the frequency response remains level but the noise spectrum falls away. This arises because the noise sidebands undergo power addition at baseband whereas the signal sidebands undergo voltage addition. Therefore the noise spectrum falls away from the baseband frequency response by 3 dB in the vicinity of zero frequency, independent of the exact shape of the i.f. filter response, i.e. even if the baseband response is not flat. The factor of 3 dB comes from the difference between power addition and voltage addition of a pair of sidebands of essentially equal level.

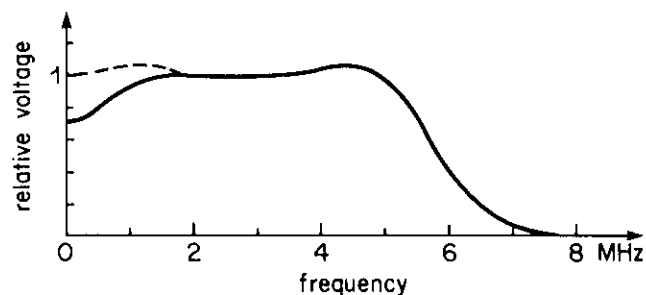


Fig. 8 - Baseband noise spectrum and overall frequency response.

--- overall response — noise spectrum

Another result of the power addition of the noise sidebands is that it is possible to obtain the noise bandwidth of the filter from the baseband noise spectrum. The relevant reference level (the 'height' of the rectangular filter) is 3 dB above the level of the spectrum near zero frequency. In practice, especially when the filter shape is non-ideal, it may be easier to determine the reference level from superimposed plots

of the baseband frequency response and the noise spectrum, as in Fig. 8. To verify the principle, the noise bandwidth was calculated from the noise spectrum and the result was 5.32 MHz. This agrees well (within 0.1 dB) with the result obtained from the actual frequency response of the filter.

The advantage of obtaining the noise bandwidth from the baseband noise spectrum is that it avoids making measurements on the i.f. stages of a receiver, which might present practical difficulties. In fact if the receiver has a video output, the need for access to the receiver circuits is entirely avoided. In addition, any baseband filtering in the receiver is taken into account.

4.2 The characteristics of a professional receiver

The second experiment was performed to find how closely the performance of a professional receiver approaches the ideal. A 'real' receiver is considerably more complex than the configuration of the first experiment; in particular there is a high gain with automatic gain control (a.g.c.), there are several filters, there are at least two frequency conversions, and a correctly-phased carrier must be derived for the synchronous demodulator.

The block diagram of Fig. 9 shows how the equipment was arranged. In the experiment, the picture signal-to-noise ratio was measured as a function of the level of the u.h.f. signal.

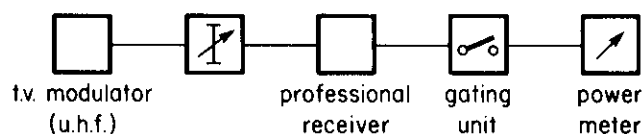


Fig. 9 - Block diagram of second experiment.

Initially, the a.g.c. characteristic of the receiver was examined. The peak-to-peak picture level at the video output was found to be 0.7 V, over a wide range of input levels, down to the lowest usable level. The a.g.c. was unaffected by picture content. Taking advantage of these characteristics, it was possible to simplify the experimental procedure.

The signal power available to the receiver was measured on a plain carrier at the peak sync level, with the attenuator at minimum. The attenuator was calibrated in a separate experiment with a resolution of 0.1 dB.

All measurements of S/N were made on a black-and-syncs video signal. The sync pulses were gated out by a gating unit so that the noise could be

measured on a power meter. The frequency response of the gating unit was flat to within 0.1 dB up to 6 MHz. The gating unit was calibrated by adding white noise, low-pass filtered to 5.5 MHz, to a video signal. The noise power was measured in the absence of the video signal, the combined signals were passed through the gating unit and the output power was measured. This was repeated at various levels over the range of signal-to-noise ratios to be measured.

The relationship between S/N and input power is shown in Fig. 10. The linearity of the relationship confirms the good performance of the a.g.c. noted earlier. It also suggests that the noise comes from the front end of the receiver. From Fig. 10, $S/N \text{ (dB)} = C \text{ (dBm)} + 78.56 \text{ dB}$.

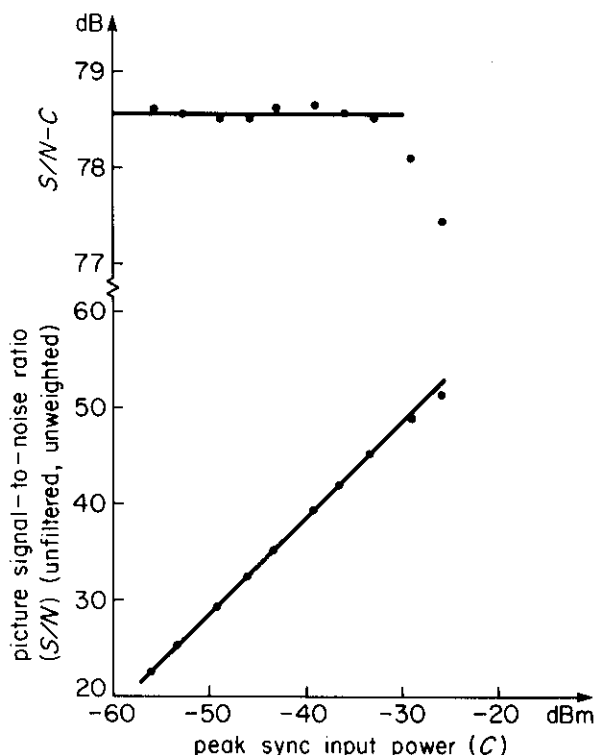


Fig. 10 - Relationship between S/N and input power for professional receiver.

The baseband noise spectrum and the frequency response are shown in Fig. 11 and these results are close to the ideal. From the noise spectrum the noise bandwidth of the receiver was found to be 5.08 MHz or 67.06 dBHz.

The noise factor of the receiver was measured by adding white noise to the signal, from a calibrated source, until the noise level at the receiver output was doubled. The signal was needed to keep the receiver's gain constant. From Fig. 10, the choice of signal level was not critical. The noise source included a band pass filter (a 'channel filter') of known noise bandwidth so that the source could be calibrated by

connecting the output to a power meter. This gave the noise power spectral density at the peak of the filter response, given that the noise spectrum incident on the filter was sufficiently flat. The noise spectrum at the output of the filter was flat to within 0.3 dB over the u.h.f. channel. The noise factor of the receiver was 19.84 dB.

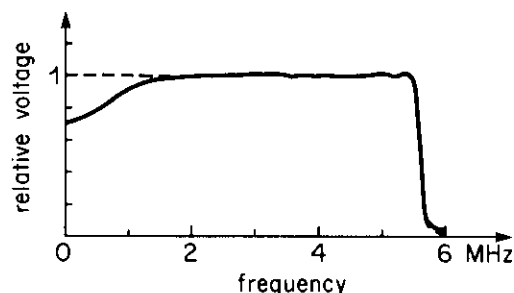


Fig. 11 - Baseband noise spectrum and frequency response of professional receiver.

--- frequency response — noise spectrum

The equivalent noise level at the input of the receiver can now be calculated. The calculation is simplified because the noise temperature of the source was determined by an attenuator at room temperature, which, to a good approximation, was equal to the standard reference temperature for noise calculations of 290 K.

Boltzmann's constant	-228.60 dBK^{-1}
Reference temperature	24.62 dBK
Noise factor	19.84 dB
Noise bandwidth	67.06 dBHz
Equivalent input noise power	-117.08 dBW or -87.08 dBm

Hence $C/N \text{ (dB)} = C \text{ (dBm)} + 87.08 \text{ dB}$

Thus $S/N = C/N - 8.52 \text{ dB}$.

This result from the professional receiver is in good agreement with theory, bearing in mind the number of additional sources of error in the second experiment.

4.3 The characteristics of a typical domestic receiver

A domestic receiver may not approach the ideal as closely as a professional receiver. In particular, the filter response is likely to be less satisfactory, making the reference levels difficult to judge. However the receiver plays an important part in determining the picture quality and in particular, the characteristics of domestic receivers are highly relevant.

Therefore a domestic receiver of recent design was also tested in the same way as the professional receiver. The domestic receiver had a surface-acoustic wave (SAW) i.f. filter and a synchronous demodulator and it was equipped with a video output. The main difference from the block diagram of Fig. 9 was that an amplifier was used to bring the video signal to standard level. The receiver was fed from a well-matched 50 ohm source throughout the tests.

The receiver was found to have a.g.c. characteristics very similar to those of the professional receiver. The linearity of the receiver circuits was checked with a staircase waveform and found to be good.

The relationship between S/N and input power is shown in Fig. 12. The range of input levels was restricted by the onset of crushing of the sync pulses. It may be noted that distortion would have set in at a lower input level if the sound carrier or adjacent-channel signals had been present. From Fig. 12, $S/N \text{ (dB)} = C \text{ (dBm)} + 89.09 \text{ dB}$.

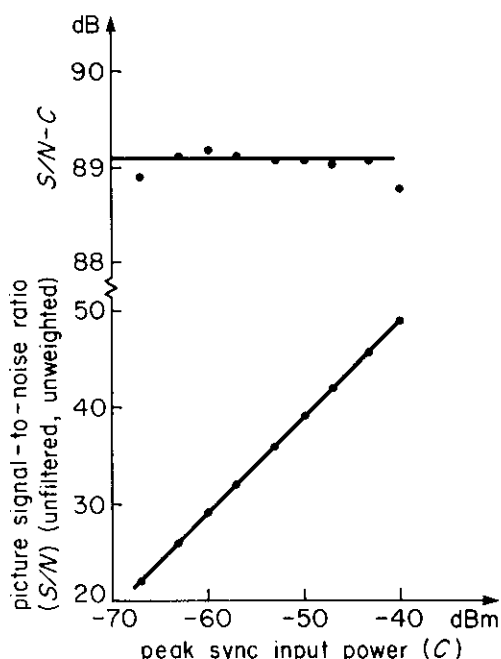


Fig. 12 - Relationship between S/N and input power for domestic receiver.

The baseband noise spectrum and the frequency response are shown in Fig. 13. The noise bandwidth was 7.22 MHz or 68.59 dBHz. From Fig. 13 we can deduce that the width of the vestigial sideband in the receiver is about 0.75 MHz.

The noise factor of the receiver was measured as 8.74 dB in the 50 ohm system. With a 75 ohm source, as specified by the manufacturer of the tuner, the noise factor was found to be 9.7 dB which is just within the manufacturer's specification.

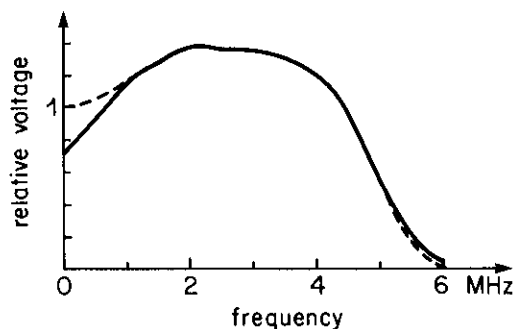


Fig. 13 - Baseband noise spectrum and frequency response of domestic receiver.

--- frequency response — noise spectrum

The equivalent input noise power can now be calculated.

Boltzmann's constant	-228.60 dBK^{-1}
Reference temperature	24.62 dBK
Noise factor	8.74 dB
Noise bandwidth	68.59 dBHz
	-126.65 dBW
	or -96.65 dBm

$$\text{Hence } (C/N)_{\text{rec}} \text{ (dB)} = C \text{ (dBm)} + 96.65 \text{ dB}$$

$$\text{and } S/N = (C/N)_{\text{rec}} - 7.56 \text{ dB.}$$

This is again in good agreement with theory.

5. SUBJECTIVE EFFECT OF THE NOISE DISPLAYED ON A DOMESTIC RECEIVER

The subjective visibility of the noise depends on the shape of the noise spectrum and also on the characteristics of the display of the receiver. In particular, the saturation control and the display tube need to be considered.

The reduction in the noise power spectral density at low frequencies will tend to make the noise less visible than noise of the same total power distributed uniformly between 0 and 5.5 MHz. On the other hand, the curtailment of the noise spectrum at high frequencies will tend to make the noise more visible than white noise.

The horizontal resolution of a colour display tube depends on the spacing of the phosphor dots or stripes. If this restricts the frequency response, noise will be less visible, but if aliasing occurs within the video bandwidth, noise will be made more visible than if the response was simply curtailed.

The noise is displayed directly as luminance noise, and also as chrominance noise through the

action of the PAL decoder. Weighting curves are available to allow the subjective effects of each type of noise, and the overall effect, to be estimated. However the visibility of chrominance noise depends not only on the shape of the noise spectrum, but also on the action of automatic chrominance control and on the setting of the viewer's saturation control (and, to a lesser extent, on the bandwidth of the chrominance circuits). If the nominal value of chrominance-luminance ratio is presented to the receiver, and if the saturation is correctly adjusted, the fall-off which is found in the video frequency response of many domestic receivers will be compensated, as far as chrominance signals and chrominance noise are concerned. The response shown in Fig. 13 of the domestic receiver used in the tests is better than many, in that the level at the colour subcarrier frequency is the same as that at low frequencies.

Subjective test results, for the visibility of the noise appearing on domestic receivers, have not been formally reported. On the other hand, the visibility of white noise displayed on a professional monitor has been investigated on many occasions. Some tests have now been carried out with v.s.b.-filtered noise also displayed on a professional monitor⁵. For the purpose of the discussion which follows it will be assumed that the result would be similar with domestic receivers. An extract from the results relevant to this study is given in Appendix 1. The tests were intended to give an up-to-date assessment, and an impairment corresponding to the limit of service (Grade 3.5) was obtained when the noise level was about 2.5 dB lower than originally assumed. It may be noted that with white noise the noise level had to be reduced by a further 2 dB.

With this information, it is possible to review the CCIR planning limits for field strength^{4,6}. In Appendix 1, it is shown that the limits of CCIR Rec. 417 are obtained if certain assumptions are made about the parameters of the receiving installation and the limiting picture signal-to-noise ratio. The assumptions are, by-and-large, still appropriate in the context of today's technology, although on the one hand the observer is about 2.5 dB more sensitive to noise, and on the other hand receiver noise factors are 1 to 2 dB better. Thus an ordinary receiving installation will give acceptable picture quality at the field strength limit of CCIR Rec. 417. Parameters for a more elaborate domestic receiving installation are also put forward in Appendix 1 and it is shown that this can give acceptable pictures at the field strength limits of CCIR Rep. 409. In this case the noise factor is determined by a masthead pre-amplifier and lower receiver noise factors give no advantage. Thus in some cases the impairment from noise will be about half a grade worse than the usual limit. However, this may be

acceptable for a small number of viewers at the fringe of a service area.

6. CONCLUSIONS AND RECOMMENDATIONS

The relationship between the vision carrier-to-noise ratio and the picture signal-to-noise ratio has been examined both in theory and by experiment. It is concluded that in System I

$$S/N = C/N - 8 \text{ dB}$$

where N is determined by the *noise bandwidth* of the receiver and S/N is unweighted.

A definition of noise bandwidth has been drawn up which is appropriate to a vestigial-sideband receiver filter; the proposed definition (given in Section 2) makes the 8 dB difference between C/N and S/N virtually independent of the receiver. An adjustment can be made to refer to the ideal r.f. noise bandwidth of 5.08 MHz, as shown in Section 3.3. The correction for a 5 MHz low pass filter in a video noise meter, if the receiver filtering is ideal, is about 0.5 dB; i.e. the -8 dB factor above becomes about -7.5 dB.

The relationship between S/N and the field strength has been derived from the parameters of the receiving installation. The characteristics of a typical domestic receiver have been presented and the implications for the visibility of noise discussed. Tests have recently been carried out, to determine the subjective visibility of both v.s.b.-filtered noise and white noise on pictures displayed on a professional monitor, and some of the results are given in an Appendix.

The characteristics of a domestic receiving installation have been quantified and it is confirmed that the CCIR planning limits for field strength of Rec. 417 and the lower limits of Rep. 409 should be maintained. The recent tests have shown an increased impairment arising from noise, compared with previous assumptions; two possible reasons for the increase are present-day expectations of picture quality, and the performance of present-day displays. However the increase is at least partially compensated by an improvement in receiver noise factors, compared with earlier assumptions.

If it is desired to confirm that the impairment caused by the noise generated and displayed by a domestic receiver is similar to the impairment caused by v.s.b.-filtered noise displayed on a professional monitor, it will be necessary to carry out subjective tests with a representative selection of domestic receivers of recent design.

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APPENDIX 1

The Connection between Field Strength and Picture Quality

As noted in the Introduction, the receiver performance is only one component in the chain linking field strength to picture quality. This Appendix examines the other links: the aerial system is considered; a limiting value of signal-to-noise ratio is postulated (i.e. a value which would define the limit of a transmitter's service area), from which it is shown how the CCIR field strength limits⁴ may be derived; finally an attempt is made to quantify the relationship between signal-to-noise ratio and picture quality.

A1.1 The power available from the aerial system

The power available from the aerial system depends upon the field strength, the gain of the aerial and the loss of the feeder. The gain of a u.h.f. aerial is usually quoted relative to that of a dipole, so we begin by considering an idealised half-wavelength dipole. (In the following we assume linearly-polarised plane waves in free space and we assume that the aerial is optimally oriented and correctly loaded.)

The power available from a half-wave dipole can be derived from the effective length, which is λ/π where λ is the wavelength^{7a,8a}. Thus the (open-circuit) voltage induced in the dipole is

$$V = E \frac{\lambda}{\pi}$$

where E is the electric field strength of the incident wave and r.m.s. values are understood.

The power available is

$$P = \frac{1}{R_a} \left(\frac{E\lambda}{2\pi} \right)^2 \quad (\text{A-1})$$

where R_a is the resistance of the idealised dipole.

It may be shown that the gain of the dipole relative to an isotropic aerial, G_a , and the resistance are related via fundamental constants^{7b} (note there is a misprint in the reference):

$$G_a = \frac{Z_o}{\pi R_a} \quad (\text{A-2})$$

where Z_o is the characteristic impedance of free space. (The power available can also be derived from the effective aperture, which is $G_a \lambda^2 / 4\pi$ ^{7c}, using (A-2) and the relationship $P_o = E^2 / Z_o$ where P_o is the power flux density of the incident wave.)

The gain of the dipole relative to an isotropic aerial can be obtained by integrating the idealised pattern. (For a method of doing this, see Ref. 8b.) The result is

$$G_a = \frac{4}{\text{Cin}(2\pi)}$$

using the notation of Ref. 9a for cosine integrals:

$$\text{Cin}(y) = \int_0^y \frac{1 - \cos t}{t} dt$$

This function is tabulated^{9b} and we find that the gain of the dipole is

$$G_a \approx 1.64$$

Hence the resistance of the dipole is

$$R_a \approx 73.1 \text{ ohms from (A-2).}$$

Thus the equation for the power available from an idealised half-wave dipole (A-1) can be expressed in decibels

$$P(\text{dBm}) = E(\text{dB } \mu\text{V/m}) + 20\log \lambda - 124.6 \text{ dB.}$$

We can now write down the power available from the aerial system. If G is the gain of the receiving aerial in decibels relative to a dipole, and L is the feeder loss (in dB; $L > 0$), then

$$C(\text{dBm}) = E(\text{dB } \mu\text{V/m}) + 20\log \lambda + G - L - 124.6 \text{ dB} \quad (\text{A-3})$$

Note: The source e.m.f. of the receiving aerial can be calculated if the system impedance, R , is known. The e.m.f. is

$$V = E \frac{\lambda}{\pi} \sqrt{\frac{R}{R_a}} \sqrt{g}$$

where $g = \log_{10}^{-1}(G/10)$

Commonly, the system impedance is 50 ohms in professional equipment and 75 ohms in domestic equipment.

A1.2 Derivations of signal-to-noise ratio and CCIR field strength limits

We begin by obtaining the relationship between the signal-to-noise ratio and the field strength.

As indicated in Section 4, the carrier-to-noise ratio can be related to the signal power at the receiver input and the parameters of the receiving installation:

$$C/N(\text{dB}) = C(\text{dBm}) - 10\log k_B T_o BF - 30 \text{ dB} \quad (\text{A-4})$$

where k_B = Boltzmann's constant ($1.38 \times 10^{-23} \text{ JK}^{-1}$);

T_o = reference temperature (290.K);

B = noise bandwidth of receiver;

F = system noise factor referred to receiver input.

The system noise factor depends on the noise temperature of the aerial (T_A), the loss of the feeder and the noise factor of the receiver (F_R).

$$F = \frac{1}{l} \left(\frac{T_A}{T_o} - 1 \right) + F_R$$

where $l = \log_{10}^{-1}(L/10)$ and the feeder is assumed to be at temperature T_o . The right-hand side of the equation can be split into three terms showing the noise contributions arising respectively from the aerial, the feeder and the receiver:

$$F = \frac{T_A}{lT_o} + \frac{l-1}{l} + (F_R - 1).$$

In practice, for a domestic u.h.f. receiving installation, the system noise factor will be approximately equal to the receiver noise factor.

From Sections 3 and 4, the signal-to-noise ratio is related to the carrier-to-noise ratio:

$$S/N \text{ (dB)} = C/N \text{ (dB)} - 8 \text{ dB} \quad (\text{A-5})$$

We can relate the signal-to-noise ratio to the field strength by combining equations (A-3), (A-4) and (A-5):

$$S/N \text{ (dB)} = E \text{ (dB } \mu\text{V/m)} + 20 \log \lambda + G - L - 10 \log k_B T_o B F - 162.6 \text{ dB} \quad (\text{A-6})$$

It is now possible to give a derivation of the CCIR field strength limits. These limits are described as the minimum field strengths for which protection may be sought, but it is reasonable to assume that the limits were determined from the minimum signal-to-noise ratio that was deemed to constitute a service. In the following, the approach and the numerical values are based on Ref. 10.

Let us postulate a signal-to-noise ratio limit of 30 dB. This applies specifically to the shape of noise spectrum produced by a receiver, with the noise being measured unweighted, through a 5 MHz low pass filter. To allow for the filter, we can add a correction factor to equation (A-6) as indicated in Section 3, or we can put $B = 4.58 \text{ MHz}$ (assuming idealised filtering). The other parameters in equation (A-6) differ between Band IV and Band V and representative values are given in Table A.1. The figures for aerial gain are not the actual gain of the aerial, but the effective gain that is likely to be realised, allowing for multipath propagation and the fact that the positioning of the aerial is a compromise over three or four channels (i.e. This is where we allow for non free-space conditions).

Table A.1: Parameters of domestic receiving installation

	Band IV	Band V
Aerial gain, dBd	9	11
Feeder loss, dB	3	4
Noise factor, dB	8	10
λ , m	0.55 (Ch. 30)	0.36 (Ch. 66)

Substituting the values from Table A.1 into equation (A-6), we find that the minimum field strength is 62.5 dB $\mu\text{V/m}$ in Band IV and 67 dB $\mu\text{V/m}$ in Band V. However, field strength contours are usually plotted in terms of median values, but it is desirable that at least 70% of locations near the edge of a service area should be served. This can be allowed for by adding a location variation factor which, from Ref. 10, is 2.5 dB for Band IV and 3 dB for Band V. This gives the CCIR field strength limits:

65 dB $\mu\text{V/m}$ in Band IV;

70 dB $\mu\text{V/m}$ in Band V.

It is possible to obtain the same signal-to-noise ratio at a considerably lower field strength. A high gain aerial realising a gain of 13 — 14 dB could be used, with a low-noise masthead preamplifier having a noise factor of 2 dB. This probably represents the limit of what is practical for a domestic receiving installation. The minimum field strength would then be 52 dB $\mu\text{V/m}$ in Band IV and 55 dB $\mu\text{V/m}$ in Band V. Thus a service area planned to the higher field-strength limits can, in effect, be extended. However, reception in an extended service area may not always be satisfactory, for two reasons. Firstly, a signal whose field strength is so low may not be adequately protected against interference. Secondly, the adverse propagation conditions that give rise to the further reduction in field strength may also give rise to severe fluctuations of field strength, and selective fading.

A1.3 The relationship between signal-to-noise ratio and picture quality

The subjective effect of noise displayed on a domestic receiver may be different from the effect of v.s.b.-filtered noise displayed on a professional monitor and, as indicated in Section 6, it would be desirable to carry out subjective tests with domestic receivers. Nevertheless results from monitors may be used as a guide.

Subjective tests have been carried out to give an assessment of the visibility of noise, based on current designs of picture monitor and in the light of current expectations of picture quality. Both white noise (bandlimited to 5.5 MHz) and v.s.b.-filtered noise (bandlimited to 5.5 MHz) were tested⁵.

An extract from the results of the subjective tests is shown in Fig. A.1, for the EBU test slide 'Girl with Toys.' On the basis of Fig. A.1, the signal-to-noise ratio limit that would be adopted now is about 32.5 dB (using a criterion of a mean grade of 3.5), compared with the value put forward in 1959 of 30 dB.

It can be seen that to give equal subjective effects, S/N for v.s.b.-filtered noise is about 2 dB less than for white noise. The difference that would be predicted by application of the hypothetical composite weighting network is only 0.6 dB and this shows that weighting curves are, at best, only approximate. It may be noted that the decoder in the tests was operated in the delay-line PAL mode; the subjective effect on the visibility of chrominance noise, relative to the simple PAL mode, is equivalent to a reduction in the chrominance noise level of about 1.5 dB.

Therefore it may be tentatively concluded that modern expectations of picture quality indicate a slightly increased limiting picture signal-to-noise ratio requirement of about 32.5 dB unweighted (measured in 5 MHz bandwidth) for VSB AM PAL System I. However, further tests, including the use of modern domestic receivers, are necessary to validate this conclusion.

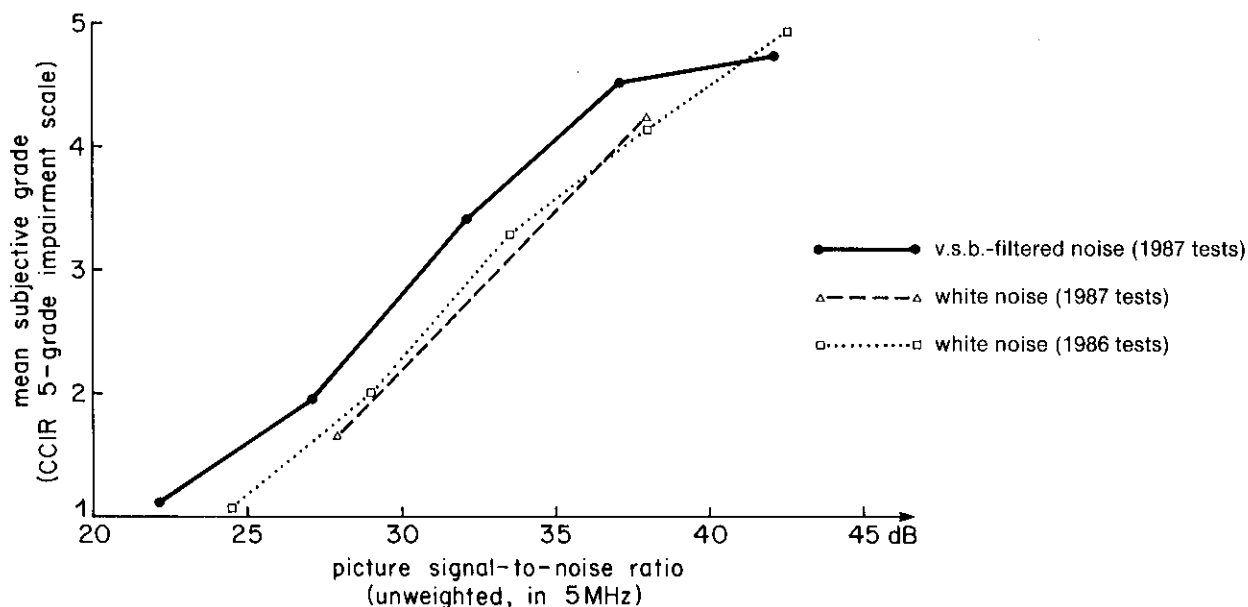


Fig. A.1 - Impairment to picture 'Girl with Toys' as a function of signal-to-noise ratio (non-expert observers).

APPENDIX 2

List of Symbols

B	noise bandwidth of filter or receiver, Hz
C	$\left\{ \begin{array}{l} \text{peak sync power at the receiver input} \\ \text{(peak sync) power available from the aerial system} \end{array} \right.$
C/N	vision carrier-to-noise ratio, power ratio or dB
$(C/N)_o$	C/N in a noise bandwidth of 5.08 MHz
$(C/N)_{rec}$	C/N in the noise bandwidth of a particular receiver
f	frequency, Hz
$H(f)$	(complex) amplitude-frequency response of filter
R	system impedance
S/N	picture signal-to-noise ratio (unweighted), voltage ratio or dB

With reference to Section 3 and Fig. 3:

Δ	a small quantity representing the skew of the filter response in the vicinity of the vision carrier frequency
C	peak sync carrier power incident on the filter
C_{IF}	peak sync carrier power emerging from the filter
G	voltage gain of demodulator
N_{BB}	r.m.s. noise voltage after demodulator
N_{IF}	noise power emerging from the filter
N_o	noise power spectral density incident on the filter
V_{BB}	r.m.s. voltage of full amplitude low frequency video tone after the demodulator
V_C	peak sync carrier voltage (r.m.s.) after the filter
V_{IF}	r.m.s. noise voltage after the filter
V_L	r.m.s. voltage of lower side tone after the filter corresponding to a full amplitude low frequency video tone
V_{NL}	noise voltage in lower sideband after the filter
V_{NU}	noise voltage in upper sideband after the filter
V_{PP}	peak-to-peak voltage of full amplitude low frequency video tone after the demodulator, i.e. peak-to-peak picture voltage
V_U	r.m.s. voltage of upper side tone after the filter corresponding to a full amplitude low frequency video tone
x	increase in S/N arising from introduction of 5 MHz low pass filter, dB

With reference to Section 3.4 and Appendix 1:

λ	wavelength, m
$\text{Cin}(\gamma)$	'entire' cosine integral
E	r.m.s. electric field strength (peak sync where appropriate)
F	system noise factor referred to receiver input
F_R	noise factor of receiver
g	gain of aerial relative to a dipole, power ratio
G	gain of aerial relative to a dipole, dB
G_a	gain of dipole relative to an isotropic aerial
k_B	Boltzmann's constant
l	feeder loss, power ratio (> 1)
L	feeder loss, dB (> 0)
P	power available from dipole (peak sync where appropriate)
P_o	power flux density (peak sync where appropriate)
R_a	resistance of idealised half-wave dipole
T_A	noise temperature of aerial
T_o	reference temperature of noise calculations
V	r.m.s. voltage induced in aerial (peak sync where appropriate)
Z_o	characteristic impedance of free space